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MEMORANDUM REPORT NO. 1701

AN ANALYTICAL APPROACH TO  
TRANSISTOR APPLICATION

by

Keats Pullen, Jr.

August 1965

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Ballistic Measurements Laboratory

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KPullen,Jr./blw  
Aberdeen Proving Ground, Md.  
August 1965

AN ANALYTICAL APPROACH TO TRANSISTOR APPLICATION

ABSTRACT

The solid-state equations defining semiconductor action in a transistor are shown to provide an analytic basis for the calculation of the behavior of a transistor in its circuit. Input and output relations and transfer relations are developed from these equations, and data on relative reliability of existing methods are developed. An example is presented which shows the ease and simplicity with which circuit design problems may be solved based on the approach.

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## INTRODUCTION

The basic equations for the current flow in a junction or a set of junctions in a solid-state device take the form of exponentials, with the exponent the product of the Fermi potential,  $\Lambda$ , and a voltage. Curiously, the true importance of these basic equations in the operation of a transistor circuit has not been generally recognized or appreciated. For that reason, it is the purpose of this report to develop, based on these equations, an analytic approach to device application. This approach serves to shed light on the practical behavior of transistors in circuits, and also yields a philosophy of operation which can easily be shown to lead to increased reliability.

## MATHEMATICAL CONSIDERATIONS

The equations for the currents flowing in the emitter and the collector for a transistor may be developed directly from the equation of action for the individual junction by recognizing that the effect of placing two junctions in close proximity to one-another is to introduce an interaction term into the simple (approximate) equation:

$$i = I_o \exp(-\Lambda V) \quad (1)$$

where  $I_o$  is the "saturation" current, and  $i$  is the current at the junction voltage,  $V$ . As given by Lo, Endres, et al, "Transistor Electronics",\* the currents in a transistor are: (Lo and others, Equation 7-1)

$$\begin{aligned} I_b &= \sigma(G)/\Lambda - \{(G_{11} + G_{21})/\Lambda\} \exp[\Lambda(v_b - I_b r_b')] \\ &\quad - \{(G_{12} + G_{22})/\Lambda\} \exp[\Lambda(v_c + I_b r_b')] \end{aligned} \quad (2)$$

$$\begin{aligned} I_c &= - \{(G_{21} + G_{22})/\Lambda\} + [\exp(\Lambda I_b r_b')] \{G_{21} \exp(\Lambda v_b) \\ &\quad + G_{22} \exp(\Lambda v_c)\}/\Lambda \end{aligned}$$

where the substitution has been made  $G_{11} + G_{12} + G_{21} + G_{22} = \sigma(G)$ . The  $G$  functions are defined by Lo et al, Equations (3) through (26).

\* Lo, A. W.; Endres, R. O.; et al. *Transistor Electronics*. Prentice-Hall Inc. Englewood Cliffs, 1955.

In the development of an analytic approach to transistor utilization, one must first examine Equations (2) to determine their meaning. For this, it is convenient first to neglect the effect of the base-spreading resistance by assuming it to be zero, and to take the initial terms to be constant, viz.,  $I_{bo}$  and  $I_{co}$ . The Equations (2) then simplify to the form:

$$\begin{aligned} I_b &= I_{bo} - \{(G_{11} + G_{21})/\Lambda\} \exp(\Lambda V_b) - \{(G_{12} + G_{22})/\Lambda\} \exp(\Lambda V_c) \\ I_c &= I_{co} + (G_{21}/\Lambda) \exp(\Lambda V_b) + (G_{22}/\Lambda) \exp(\Lambda V_c). \end{aligned} \quad (2a)$$

The correction for non-zero values of the base-spreading resistance is easily made where required.

Now, it should be noted that all of the factors,  $G_{11}$ ,  $G_{12}$ ,  $G_{21}$ ,  $G_{22}$ ,  $\sigma(G)$ , and  $\Lambda$  are physical parameters which are dependent on the basic properties of the semiconductor material itself. The Fermi potential is even more fundamental than the other factors, in that it depends for practical purposes only on the absolute temperature.

If one assumes, again, that the values of  $I_{bo}$  and  $I_{co}$  can be neglected, (this is largely true, since in practice we forward-bias the device from a condition of cutoff, and the leakage current is assumed negligible compared to the operating currents) the ratio of the collector to the base current may be established. Taking the terms involving  $V_c$  also as negligible (with typical devices this is the meaning of the relative constancy of  $I_b$  and  $I_c$  with variation of the collector voltage,  $V_c$ ) the ratio may be written as:

$$I_c/I_b = G_{21}/(G_{11} + G_{21}) \gg 1. \quad (3)$$

Examination of Equation (3) leads to a very important conclusion. The ratio indicated can only be achieved if  $|G_{21}| \approx |G_{11}| \pm \Delta$ , where  $\Delta$  is small compared to either  $G_{11}$  or  $G_{21}$  magnitude-wise, and the signs of  $G_{11}$  and  $G_{21}$  of necessity must differ.

It is now evident why the current gain in a transistor is rather difficult to stabilize from device to device or over a given operating range on a given device. Clearly, the achievement of a high value of

current gain is only possible when the magnitudes of  $G_{11}$  and  $G_{21}$  are almost identical and their signs differ. Only a very small change in the value of either of these parameters, resulting from such things as processing variations, can lead to considerable change in the value of the ratio of  $I_c/I_b$ . Actual experience amply supports the analytic conclusion.

Since there are three components for each of the currents  $I_b$  and  $I_c$ , it is convenient to rewrite Equations (2) in the form:

$$\begin{aligned} I_b &= I_{bo} + I_{bl} + I_{b2} \\ I_c &= I_{co} + I_{cl} + I_{c2} \end{aligned} \quad (4)$$

where the subscripts 1 and 2 refer to the base voltage terms and the collector voltage terms respectively. These sub-elements are useful in the expression of the derivatives as will be seen in a later paragraph.

Actually, all kinds of active devices, transistors included, are dependent for their functioning not only on a static environment, but also a small-signal environment. The static environment is dependent on voltages  $V_b$  and  $V_c$  and currents  $I_b$  and  $I_c$ , and the small-signal environment depends on  $V_b$  and  $I_c$  in particular. Since the equations for the currents depend directly on the voltages, as can be seen in Equations (2a), the small-signal environment must be dependent on partial derivatives of currents with respect to the two voltage variables. The resulting admittances (or conductances) are:

$$\begin{aligned} y_i &= \partial I_b / \partial V_b = - (G_{11} + G_{21}) \exp (\Lambda V_b) = \Lambda I_{bl} = y_i \\ y_f &= \partial I_c / \partial V_b = G_{21} \exp (\Lambda V_b) = \Lambda I_{cl} = y_f \\ y_r &= \partial I_b / \partial V_c = - (G_{12} + G_{22}) \exp (\Lambda V_c) = \Lambda I_{b2} = y_r \\ y_o &= \partial I_c / \partial V_c = G_{22} \exp (\Lambda V_c) = \Lambda I_{c2} = y_o \end{aligned} \quad (5)$$

These equations show that all of the conductances for a transistor are dependent on the static current level for the appropriate component of current and on a physical constant, the Fermi potential.

Of these currents, only two ordinarily are of significant magnitude, namely,  $I_{bl}$  and  $I_{cl}$ . And, as has been already noted, the values of  $I_c$  and  $I_b$ , can not both be specified for any given transistor device. This is also true of  $I_{bl}$  and  $I_{cl}$ . Hence, a selection must be made to specify one or the other.

A similar discussion to that for  $I_c/I_b$  can be used to show that usually

$$|G_{22}/(G_{12} + G_{22})| > 1 \quad (6)$$

and consequently, that  $|G_{22}| \approx |G_{12}|$  to within a very small increment, which in this instance need not be as small as with Equation (3).

The value of the Fermi potential of course is dependent on the absolute temperature, and as a consequence can vary with operating conditions. It decreases with an increase in temperature. As a result, design for use at somewhat elevated temperatures should be based on a value of the potential of approximately 35,000 micromhos per millampere rather than the 39,000 used at room temperature. Since the range of variation can be specified rather precisely, compensation for the variation is not difficult to accomplish.

#### STABILIZATION

Now, it is generally known that the relative values of the current components for good transistors must satisfy the inequality:

$$I_{cl} > I_{bl} \gg I_{c2} \stackrel{>}{\approx} I_{b2} \quad (7)$$

In actual operation, it is convenient to stabilize certain variables for the active device in order to assure that the correct operating environment can be maintained. Normally, since the collector supply voltage is set, it is necessary to stabilize the static collector current in order to assure that the correct range of operating conditions will be available. (This is the reason that cathode bias techniques have proven so valuable with electron tubes.)

Because of the instability of the ratio in Equation (3), it is clearly not possible to get a high degree of stabilization of collector current based on the stabilization of base current, even though the value of collector current is less sensitive to temperature change with stabilized base current than it would be if  $V_b$  alone were stabilized. For this reason, best stabilization is obtained by regulating the value of the emitter current ( $I_b + I_c$ ), rather than the base current, through the use of an emitter resistance and a controlled base voltage. This arrangement is the equivalent of the cathode-bias regulation used with tube circuits. Since usually  $I_b \ll I_c$ , stabilizing  $I_e$  accomplishes the required stabilization of  $I_c$ .

By stabilizing the value of  $I_c$  through emitter degeneration, one obtains a known and controlled variation of the transadmittance as a function of the collector current through the equation:

$$y_f = \Lambda I_{cl} . \quad (5b)$$

Clearly, the value of  $\Lambda$ , the Fermi constant, is known and is dependent on the electronic charge, Boltzmann's constant, and the absolute temperature, and its value in the temperature range of interest is between 35000 and 39000 micromhos per milliampere. If Equation (5b) is used as the basis for establishing the operating characteristics for a transistor in a circuit, and then the circuit is so designed that the power required by the input circuit does not degrade the operation of the source circuit, stable operation is an automatic consequence. For this condition to exist, it is necessary that the source impedance,  $Z_s$ , of the input signal generator for the amplifier satisfy the relation:

$$Z_s < 1/y_i \quad (8)$$

where  $y_i$  is the input admittance of the amplifier. This condition can easily be satisfied in several ways, the easiest being the use of a transistor operated as an emitter follower. Subject to some compound load-line restrictions, the output impedance of an emitter follower is given by the equation:

$$Z_o = 1/(y_{i'} + y_{f'}) \quad (9)$$

which can clearly be made smaller than  $1/y_i$ . Consequently, a stable transistor amplifier will often include pairs consisting of an impedance transformer and an amplifier, each of which is based on the use of transistors. The active transformer stages and the amplifier stages will alternate in such a configuration.

In actual practice, the input admittance includes both a conductive and a capacitive component. The conductive component is the one which is specified in terms of Equation (5a). There are two parts of the capacitive component, a diffusion component, which is also proportional to the device current, and a transition capacitance, which is a function of collector voltage. Over normal operating ranges, the diffusion capacitance predominates, but at low values of collector current and collector voltage, the transition capacitance is more important. The diffusion capacitance does not increase quite as rapidly as the transconductance increases, with the result that the values of  $f_\alpha$  and  $f_t$  both increase slowly with collector current where  $f_\alpha$  is the cut off frequency and  $f_t$  is the frequency for unity current gain.

The approximate value of the diffusion capacitance may be determined, for a given collector current, in terms of the equation:

$$2\pi f_\alpha C_d \doteq \Lambda(I_c + I_b) \doteq \Lambda I_c$$

or (10)

$$C_d \doteq \Lambda I_c / 2\pi f_\alpha$$

Since the order of magnitude of the capacitance is really what is required for circuit design purposes, either  $f_\alpha$  or  $f_t$  may be used in the equation, depending on which is available. It is usually safe to assume that the total capacitance is less than twice the value obtained from Equation (10).

In practice, the effect of the base-spreading resistance can not be ignored, either for its effect on the input impedance for the transistor or for its effect on the transfer admittance. An equation can be derived

which gives the value of the internal base signal voltage (intrinsic value), the voltage across the junction itself, in terms of the terminal voltage,  $V_b$ :

$$V_{b'} = V_b / (1 + y_i' r_b') \quad (11)$$

where  $y_i'$  is the input admittance of the intrinsic junction itself. It may be evaluated in terms of the base current and Equation (5a).

As a consequence of Equation (11), the overall transadmittance for the transistor may be given in terms of the equation:

$$y_f = y_f' \times (V_{b'}/V_b) = y_f' / (1 + y_i' r_b') . \quad (12)$$

Since at least  $y_i'$  includes both a conductance and a susceptance, the value of  $y_f$  can have a susceptance component even though  $y_f'$  may not. In addition, to make this expression reasonably complete, a time delay term should be included:

$$y_f = y_f' \exp(-s\tau) / (1 + y_i' r_b') . \quad (13)$$

Measurements made at the Ballistic Research Laboratories have shown that the time delay  $\tau$  is almost totally independent of operating conditions, voltage and current, as it should be, since it is dependent on diffusion velocity.

#### COMPARISON WITH THE HYBRID REPRESENTATION

One of the more widely used representations of a transistor utilizes the "hybrid" or "H" parameters. In this system of representation, the two principal parameters are those for the input impedance and the forward transfer ratio. Giving these in terms of the previously considered relations yields:

$$\begin{aligned} h_{ie} &= 1/y_{ie} = 1/\Lambda I_b \\ h_{fe} &= y_{fe}/y_{ie} = I_c/I_b . \end{aligned} \quad (14)$$

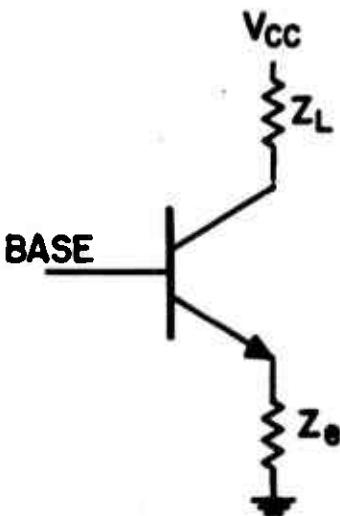
The first observation which may be made about Equations (14) is that the parameter,  $h_{le}$ , which does involve the Fermi potential, is also a function of the current  $I_b$ , which, as has been noted, cannot be stabilized. And the second is that the parameter,  $h_{fe}$ , is independent of the Fermi potential and is also dependent on  $I_b$ . Consequently, neither of these parameters can be stabilized in terms applicable in practical circuits. Stabilization can only be achieved, in fact, by either implicitly or explicitly accepting the fact that either  $y_i$ , or  $y_f$ , may be stabilized, but not both. Since the stabilization of  $y_f$ , is the more important in that it is necessary to maintain a closely-controlled static operating point as a function of collector voltage and current, the device naturally is used with a stabilized trans-admittance.

#### APPLICATION OF THE THEORY

Application of the above theory to a practical example will show how simple and straight-forward transistor circuit design can become when the above relations are utilized. Since the problem of proper circuit design is clearly one of coordinating a static with a small-signal design, a review of both the basic circuits and the basic equations to be applied is first desirable.

As has been noted above, the desired approach to a coordinated static and small-signal operation calls for stabilization of the static emitter current in the transistor. This stabilizes the total positive and negative deviation achievable in the collector circuit, and also stabilizes the transconductance variation with the applied signal. The result is a precisely controlled response for the network as a whole.

Stabilization of the emitter current is best accomplished by the stabilization of voltage difference across a fixed resistance. Typically, this is accomplished by "forward-biasing" the base of the transistor by at least one volt more than the normal forward potential, and placing a resistor of proper value in the emitter circuit to permit the desired current to flow, (Figure 1). This stabilization is a form of dc



**FIG. I - BASIC TRANSISTOR CONNECTION**

degeneration. The presence or absence of ac or signal degeneration depends on the presence or absence of bypassing by a capacitor or of a cancelling signal due to a differential element.

It should be noted that the required forward bias is not a current bias, but a fixed voltage bias, since the stabilization of emitter current through the voltage loss across the emitter resistor is only effective if the static base-to-ground voltage is stabilized. Then as long as the voltage from emitter to ground,  $V_e$ , satisfies the relation:

$$V_e \gg \Delta V_B \quad (15)$$

where  $\Delta V_B$  is the increment in the base-to-emitter voltage due to temperature and other factors, stabilization of  $I_e$  and  $I_c$  will be accomplished.

The basic equations for the input and output admittance and the forward transfer admittance for the generalized transistor amplifier are (Figure 2):

$$Y_{id} = [y_i + \Delta y'(Z_e + Z_L)] / \{1 + y_i r_b + \sigma(y') Z_e + y_o Z_L + \Delta y' [Z_e Z_L + r_b'(Z_e + Z_L)]\} \quad (16)$$

where  $\Delta y' = y_i' y_o - y_f' y_r = y_i' y_c'$ , and  $\sigma(y') = y_i' + y_f' + y_r + y_o$  as before. Similarly, the forward admittance for the amplifier is:

$$Y_{fd} = -[y_f' - \Delta y' Z_e] / \{1 + y_i' (r_b' + Z_s) + \sigma(y') Z_e + y_o Z_L + \Delta y' [Z_e Z_L + (r_b' + Z_s)(Z_e + Z_L)]\} \quad (17)$$

with the same definitions applying. The  $Z_s$  here is the input impedance introduced by the signal source. For the output admittance:

$$Y_{od} = [(y_o + \Delta y' (r_b' + Z_e + Z_s))] / \{1 + y_i' (r_b' + Z_s) + \sigma(y') Z_e + \Delta y' (r_b' + Z_s) Z_e\} \quad (18)$$

Fortunately, the terms involving  $y_o$ ,  $\Delta y'$ , and  $y_r$  frequently can be neglected, giving the following simplified forms:

$$Y_{id} = y_i' / \{1 + y_i' r_b' + \sigma(y') Z_e\} \quad (16a)$$

$$Y_{fd} = -y_f' / \{1 + y_i' (r_b' + Z_s) + \sigma(y') Z_e\} \quad (17a)$$

and

$$K_v = Y_{fd} Z_L \quad (17b)$$

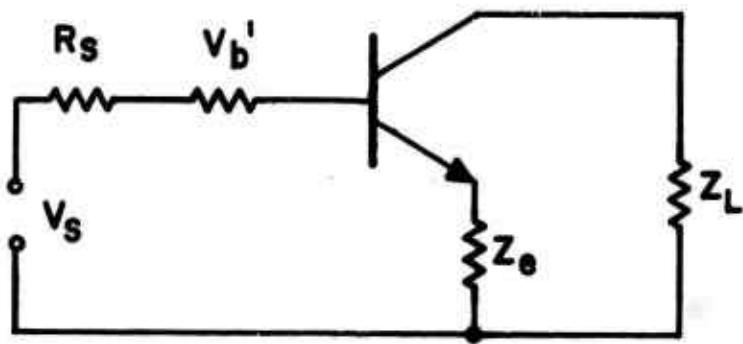
$$Y_{od} = y_o / \{1 + y_i' (r_b' + Z_s) + \sigma(y') Z_e\} \quad (18a)$$

The equations are simple both in format and in use, particularly when the  $\sigma(y')$  factor is simplified to:

$$\sigma(y') \approx y_i' + y_f' \approx y_f' \quad (19)$$

The last approximation assumes that the beta of the transistor is at least fifty or more.

Equations (16a), (17a), and (18a) show further why the stabilization of  $I_c$  has been selected in preference to stabilization of  $I_b$ . All of the largest terms involving the device parameters in the denominators of these equations involve  $y_f'$ , and the only important term in the numerator of Equation (17a) involves  $y_f'$  also; the stabilization of these three admittances is therefore best achieved through stabilizing  $I_e$ , and through it  $I_c$ , to stabilize  $y_f'$ .



**FIG. 2 - GENERAL AC CIRCUIT FOR TRANSITOR AMPLIFIER**

If the emitter impedance,  $Z_e$ , Figure 2, has the form of a parallel resistor and capacitor, its mathematical representation takes the form:

$$Z_e = R_e / (1 + j\omega C_e R_e) . \quad (20)$$

At very low frequencies, this expression simplifies to the form:

$$Z_e = R_e \quad (20a)$$

and dc stabilization results. If, however, at higher frequencies, the magnitude of the term  $\omega C_e R_e$  is large compared to unity, then the equation takes the simplified form:

$$Z_e = - j/\omega C_e \quad (20b)$$

and if the magnitude of the expression  $\sigma(y')/\omega C_e$  is small compared to unity, then high-frequency degeneration does not exist. In fact, high frequency degeneration is normally used for one of the following reasons:

- (1) to raise the input impedance by increasing the denominator of Equation (16a),
- (2) to reduce the variation of  $y_f' = \Delta I_c$  due to changes of  $I_c$ , and
- (3) to make the active device more closely approximate a current source.

These functions, of course, can be achieved by other ways as well as by the use of emitter degeneration, but they are most easily achieved by degeneration.

The discussion in this report is limited to the simple amplifier having degeneration introduced through the use of a common emitter resistor. This should not be taken to imply, however, that the analysis is in any way limited in its application to this form of circuit alone, since any transistor circuit whose behavior can be expressed in terms of the admittance parameters of the active device can equally well be studied by the described technique. As is disclosed in the writer's book "Handbook of Transistor Circuit Design"\*, there are few such circuits which offer difficulty in these terms. Oscillators, mixers, multivibrators, power amplifiers, and many other types of circuits are considered in these terms in the referenced "Handbook".

#### AN EXAMPLE

As an example of the use of this design procedure, assume that a video amplifier is required to have a voltage amplification of at least twenty over the collector current range between one and ten milliamperes. The active device chosen has a base-spreading resistance of 50 ohms, a minimum beta of 20, and a limit frequency,  $f_t$ , of 100 megacycles per second (megaHertz). An attempt is to be made to design such an amplifier, the bandwidth to extend to five megaHertz.

The intrinsic transconductances at the current limits may be read directly from Equation (5b) as a minimum of 39,000 and a maximum of 390,000 micromhos, with an average value of about 214,000 micromhos. The effective values of  $C_{be}$  may also be determined from Equation (10) to be about 65 pfd at the lower transconductance value and 650 pfd at the larger. The average value is about 360 pfd.

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\* Pullen, Keats A., Jr. *Handbook of Transistor Circuit Design*. Prentice-Hall, Inc. Englewood Cliffs, 1961.

The emitter current for the transistor may be stabilized at five milliamperes by forward-biasing the base by one volt plus the static bias required to cause the passage of five milliamperes, and by introducing 200 ohms in the emitter lead to provide the required volt of voltage loss. If the transistor is a silicon device, rather than a germanium, the forward bias on the emitter should be at least two volts if maximum stability is desired.

If the minimum value of beta for the transistor is 20, then the input conductance values at the two limits are at most 2000 and 20,000 micromhos, corresponding to input resistances in excess of 500 to 50 ohms, respectively. The equivalent input network therefore consists of a series resistance of fifty ohms value (the base-spreading resistance), and a parallel combination of a variable capacitance ranging between 65 and 650 pfd, and a variable resistance whose value ranges from "in excess of" 500 ohms to "in excess of" 50 ohms. The response of this network as a function of frequency may be determined for each limiting condition and any other points desired, and the three-db frequency for each choice calculated in conventional manner. The lowest corner frequency obtained from this set of calculations specifies the approximate upper frequency limit for the circuit. A superficial estimate indicates that the response of the network, and the amplifier therefore, will be just adequate.

As soon as the decision has been made that the required characteristics can be achieved, it is important to determine whether gain stabilization will be required or not. Based on the gain equation, Equation (17a), and neglecting  $Z_s$ , the load resistance value may be selected which will yield a stage gain of 20 at the minimum current of one milliampere. For this, the value of  $R_L$  is 500 ohms. Then the intrinsic value of  $K_v$  (neglecting  $r_b$ ) Equation (17b) may be determined at ten milliamperes to be 200; after including the effect of base-spreading resistance the value is about 100. If any kind of linearity is required, emitter degeneration or some other feedback is required, the load resistance being readjusted to keep the minimum gain at 20. This emitter resistance,

$R_e$ , may be chosen so that its effect is negligible at one milliamperes, yet it can introduce a significant limitation into the gain at higher values of current. For example introduction of six ohms into the emitter circuit will reduce the gain at minimum current by twenty percent yet at the high-current limit it will reduce the gain to three-tenths of its undegenerated value.

Typically, an emitter resistance large enough to make the value of the emitter degeneration term equal to unity, namely:

$$\text{Re}[y_f' R_e] = 1 \quad (21)$$

where  $\text{Re}$  stands for "real part of", at the minimum current and transconductance gives excellent stabilization of gain. Using this condition, and a value of  $R_L$  of 1000 ohms to compensate for the loss, the emitter resistance required is 25 ohms, and the range in amplification is from 20 to approximately 33. Clearly, this is a very significant improvement.

The linear variation of transconductance of course can be utilized where nonlinearity is actually required, and the knowledge of how it varies can be invaluable to the designer.

KEATS PULLEN, JR.

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